I/Q Imbalance Mitigation for Time-Reversal STBC Systems Over Frequency-Selective Fading Channels

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***Abstract*— This work studies the effect of in-phase/quadrature (I/Q) imbalance on time-reversal space-time block coded (TR-STBC) communication systems operating over frequency-selective fading channels. The transceiver I/Q imbalance in a 2×1 TR orthogonal STBC (TR-OSTBC) system is studied in detail, and low-complexity mitigation solutions are proposed by exploiting the special structure of the received data. Our results show that the proposed solutions both in time domain and in frequency domain can effectively mitigate the I/Q distortion in such a system.**

***Index Terms*— I/Q imbalance, time reversal, space-time block coding**

I. Introduction

STBC communication systems provide reliable data transmission by exploiting spatial diversity in flat fading channels [1]. To further exploit the multipath diversity embedded in the frequency-selective fading channels, TR-STBC system, both orthogonal and quasi-orthogonal, have been proposed and studied extensively [2]–[5].

In practical implementation, the I/Q imbalance (the non-ideal matching between the relative amplitudes and phases of I and Q branches of a transceiver) exists in many RF systems due to analogue imperfections [6]. This commonly results in a small complex conjugate term in time domain (see eqs.(1-2)), hence an equivalent mirror-image distortion term in frequency domain in the data structure of communication systems [7] [8]. Therefore, the I/Q imbalance increases symbol error rate (SER) drastically, especially in STBC systems utilizing both symbols and their complex-conjugates.

Although there exist various I/Q mitigation methods in literature, the methods in [8] [9] are preferred when the *transmitter* I/Q imbalance is taken into account. The receiver I/Q imbalance compensation is studied for an OSTBC-OFDM systems over frequency-selective fading channels in [8]. Taking into account the transmitter I/Q imbalance, the transceiver I/Q imbalance compensation for STBC OFDM systems in studied in [9]. Specifically, the method in [9] can mitigate the *transceiver* I/Q imbalance at the receiver with only the estimated effective channel state information (ECSI) ($\{\mathbf{u}_i, \mathbf{v}_i\}$ in eq.(4)). It combines the tasks of estimating the transmitter I/Q imbalance parameters $\{\alpha_T, \beta_T\}$, the channel $\mathbf{h}_i$ and receiver I/Q imbalance $\{\alpha_R, \beta_R\}$. In this work, we develop a new transmission scheme that enables simple yet effective solutions, both in time domain and in frequency domain, to mitigate transceiver I/Q imbalance for TR-OSTBC systems operating over frequency-selective fading channels. Simulation results demonstrate that the transceiver I/Q imbalance can be effectively compensated by employing the proposed solutions with either known or estimated ECSI.

Notation: $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ denote complex-conjugate, transpose and complex-conjugate transpose operations, respectively; $\Re\{\cdot\}$ and $\Im\{\cdot\}$ denote real and imaginary parts, respectively; $\circledast$ and $\otimes$ denote convolution and Kronecker product, respectively; $(\cdot)_K$ denotes mod-K; circ$(\mathbf{x})$ denotes a circulant matrix with $\mathbf{x}$ being its first column.

II. Signal and System Models

A. Transceiver I/Q Imbalance Model

In this work, we adopt a two-parameter frequency-independent I/Q imbalance model, and assume the same transmitter/receiver I/Q imbalance parameters for all transmitting/receiving antennas sharing *the same* local oscillator (LO).

When I/Q imbalance exists, the LO output at the transmitter can be expressed as $c_T(t) = \alpha_T e^{j\omega_c t} + \beta_T^* e^{-j\omega_c t}$, where $\alpha_T = \frac{1}{2}[1 + (1 + \epsilon_T)e^{j\varphi_T}]$, and $\beta_T = \frac{1}{2}[1 - (1 + \epsilon_T)e^{j\varphi_T}]$, with ϵ_T and φ_T representing amplitude and phase imbalance parameters of the transmitting antenna. Consequently, the up-converted band-pass signal for the intended transmission $s(t)$ becomes $\Re\{s(t)c_T(t)\} = \Re\{s_T(t)e^{j\omega_c t}\}$, resulting in the equivalent baseband signal

$$s_T(t) = \alpha_T s(t) + \beta_T s^*(t), \qquad (1)$$

containing both the intended signal and its complex-conjugate.

Similarly, the LO output at the receiver can be expressed as $c_R(t) = \alpha_R^* e^{-j\omega_c t} + \beta_R e^{j\omega_c t}$, where $\alpha_R = \frac{1}{2}[1 + (1 + \epsilon_R)e^{j\varphi_R}]$, and $\beta_R = \frac{1}{2}[1 - (1 + \epsilon_R)e^{j\varphi_R}]$, with ϵ_R and φ_R representing amplitude and phase imbalance parameters of the receiving antenna. Taking into account the effect of channel $h(t)$ and additive noise $n(t)$, the received data down-converted by $c_R(t)$ takes the equivalent baseband form,

$$r(t) = \alpha_R^*[h(t) \circledast s_T(t)] + \beta_R[h(t) \circledast s_T(t)]^* + n(t). \qquad (2)$$

B. Data Structure With CP Aided Transmission

When two transmit-antennas are used to simultaneously transmit $K \times 1$ data blocks $\mathbf{s}_i$ over frequency-selective fading

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channels $\mathbf{h}_i = \begin{bmatrix} h_i[0] & \cdots & h_i[L] \end{bmatrix}^T, (i = 1, 2, \; L \ll K)$, the received $K \times 1$ data vector containing I/Q imbalance has the following discrete-time baseband form,

$$\mathbf{r} = \sum_{i=1}^{2} (\mathbf{U}_i \mathbf{s}_i + \mathbf{V}_i \mathbf{s}_i^*) + \mathbf{n}. \tag{3}$$

Here the square matrices $\mathbf{U}_i$ and $\mathbf{V}_i$ in (3) are circulant due to the use of a cyclic-prefix (CP) preamble copying from L trailing symbols of the data vector $\mathbf{s}_i$ at the transmitter. Specifically, matrices $\mathbf{U}_i = \text{circ}(\mathbf{u}_i)$ and $\mathbf{V}_i = \text{circ}(\mathbf{v}_i)$ are parameterized by the channel and transceiver I/Q imbalance parameters as follows,

$$\begin{matrix} \mathbf{u}_i = \alpha_T \alpha_R^* \tilde{\mathbf{h}}_i + \beta_T \beta_R^* \tilde{\mathbf{h}}_i^* \\ \mathbf{v}_i = \beta_T^* \alpha_R^* \tilde{\mathbf{h}}_i + \alpha_T^* \beta_R^* \tilde{\mathbf{h}}_i^* \end{matrix}, \qquad i = 1, 2. \tag{4}$$

Here the $K \times 1$ vector $\tilde{\mathbf{h}}_i = \begin{bmatrix} \mathbf{h}_i^T & \mathbf{0}_{K-L+1}^T \end{bmatrix}^T$ is a zero-padded channel. Compared to the effect of I/Q imbalance on the signal, such effect on noise is relatively small at reasonably high SNRs, hence a proper $\mathbf{n} \sim \mathcal{CN}(\mathbf{0}, \sigma_n^2 \mathbf{I}_K)$ is adopted.

The matrices $\mathbf{U}_i$ and $\mathbf{V}_i$ can be diagonalized by the unitary discrete Fourier transform (DFT) matrix $\mathbf{W}$ as $\mathbf{U}_i = \mathbf{W}^H \mathbf{\Lambda}_{u_i} \mathbf{W}$, and $\mathbf{V}_i = \mathbf{W}^H \mathbf{\Lambda}_{v_i} \mathbf{W}$, with $\mathbf{\Lambda}_{u_i} = \text{diag}(\mathbf{W}\mathbf{u}_i)$, and $\mathbf{\Lambda}_{v_i} = \text{diag}(\mathbf{W}\mathbf{v}_i)$.

C. Data Model for the Proposed TR-OSTBC System With CP

In a 2×1 TR-OSTBC system, symbol vectors in the $2K \times 2$ codeword matrix

$$\mathbf{C}(\mathbf{s}_1, \mathbf{s}_2) = \begin{pmatrix} \mathbf{s}_1 & \mathbf{s}_2 \\ -J_c(\mathbf{s}_2) & J_c(\mathbf{s}_1) \end{pmatrix} \qquad \begin{matrix} \longrightarrow \text{space} \\ \downarrow \text{time} \end{matrix}$$

are transmitted over two consecutive time slots through two transmit-antennas. Here $J_c(\cdot)$ is a time-reversal conjugate (TRC) operator such that $J_c(\mathbf{s}_i) = \mathbf{J}\mathbf{s}_i^*$ with $\mathbf{J}$ being an exchange matrix. For any circulant matrix $\mathbf{A}$, we have $\mathbf{J}\mathbf{A}^*\mathbf{J} = \mathbf{A}^H$. Using the results in (3), the received data block in a TR-OSTBC system over two consecutive time slots can be formulated as $\vec{\mathbf{r}} = \mathbf{U}\vec{\mathbf{s}} + \mathbf{V}\vec{\mathbf{s}}^* + \vec{\mathbf{n}}$, or, specially,

$$\begin{bmatrix} \mathbf{r}_1 \\ \mathbf{J}\mathbf{r}_2^* \end{bmatrix} = \begin{bmatrix} \mathbf{U}_1 & \mathbf{U}_2 \\ \mathbf{U}_2^H & -\mathbf{U}_1^H \end{bmatrix} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{s}_2 \end{bmatrix} + \begin{bmatrix} \mathbf{V}_1 & \mathbf{V}_2 \\ \mathbf{V}_2^H & -\mathbf{V}_1^H \end{bmatrix} \begin{bmatrix} \mathbf{s}_1^* \\ \mathbf{s}_2^* \end{bmatrix} + \begin{bmatrix} \mathbf{n}_1 \\ \mathbf{J}\mathbf{n}_2^* \end{bmatrix}. \tag{5}$$

In the absence of I/Q imbalance, $\mathbf{V} = \mathbf{0}$, symbol blocks $\mathbf{s}_1$ and $\mathbf{s}_2$ can be decoded separately due to the circulant nature of matrices $\{\mathbf{U}_i\}_{i=1}^2$. However, the presence of I/Q imbalance forces us to take into consideration a widely linear relation containing both $\vec{\mathbf{s}}$ and its complex-conjugate $\vec{\mathbf{s}}^*$ in (5) to effectively decode symbols $\{\mathbf{s}_i\}_{i=1}^2$.

III. Proposed Solutions

A. Time Domain Processing

In the presence of I/Q imbalance, to fully capture the symbol information contained in (5), we can re-arrange data as

$$\underbrace{\begin{bmatrix} \Re\{\vec{\mathbf{r}}\} \\ \Im\{\vec{\mathbf{r}}\} \end{bmatrix}}_{\tilde{\mathbf{r}}} = \underbrace{\begin{bmatrix} \Re\{\mathbf{U}+\mathbf{V}\} & \Im\{\mathbf{V}-\mathbf{U}\} \\ \Im\{\mathbf{U}+\mathbf{V}\} & \Re\{\mathbf{U}-\mathbf{V}\} \end{bmatrix}}_{\mathbf{H}} \underbrace{\begin{bmatrix} \Re\{\vec{\mathbf{s}}\} \\ \Im\{\vec{\mathbf{s}}\} \end{bmatrix}}_{\tilde{\mathbf{s}}} + \begin{bmatrix} \Re\{\vec{\mathbf{n}}\} \\ \Im\{\vec{\mathbf{n}}\} \end{bmatrix}. \tag{6}$$

Under white Gaussian noise (WGN) assumption, the optimal maximum-likelihood sequence estimate (MLSE) of $\{\mathbf{s}_i\}_{i=1}^2$ can be obtained from

$$\hat{\tilde{\mathbf{s}}}_{\text{ML}} = \arg \min_{\tilde{\mathbf{s}} \in \mathcal{S}} \|\tilde{\mathbf{r}} - \mathbf{H}\tilde{\mathbf{s}}\|^2, \tag{7}$$

where the set $\mathcal{S}$ contains all possible $4K \times 1$ signal vectors $\tilde{\mathbf{s}}$ from a given symbol constellation. For a TR-OSTBC system with large data block size K (typically $K \gg L$), to reduce the computational complexity of the MLSE, a low-complexity sub-optimal solution built upon the filtering idea can be found for practical implementation. Introducing a linear filter $\mathbf{F}$, we can minimize $E\{\|\tilde{\mathbf{s}} - \mathbf{F}\tilde{\mathbf{r}}\|^2\}$ to obtain the linear minimum-mean-square-error (LMMSE) estimates of $\{\mathbf{s}_i\}_{i=1}^2$,

$$\hat{\tilde{\mathbf{s}}}_{\text{LMMSE}} = \mathbf{F}_{\text{LMMSE}}\tilde{\mathbf{r}} = \mathbf{H}^T(\mathbf{H}\mathbf{H}^T + \sigma_n^2/\sigma_s^2 \mathbf{I}_{4K})^{-1}\tilde{\mathbf{r}},$$

where σ_s^2 is signal power per sample. Its zero-forcing (ZF) counterpart is simply the result of dropping the diagonal loading factor,

$$\hat{\tilde{\mathbf{s}}}_{\text{ZF}} = \mathbf{F}_{\text{ZF}}\tilde{\mathbf{r}} = \mathbf{H}^T(\mathbf{H}\mathbf{H}^T)^{-1}\tilde{\mathbf{r}}.$$

Note that the inversion of the large-size ($4K \times 4K$) matrices in the above LMMSE and ZF solutions can be avoided explicitly due to the proposed CP based TR-OSTBC scheme. The block structure of the matrix $\mathbf{H}$ in (6) results in a block Gram-matrix $\mathbf{G} = \mathbf{H}\mathbf{H}^T$ with the following structure,

$$\mathbf{G} = \begin{bmatrix} \mathbf{G}_{11} & \mathbf{G}_{12} \\ \mathbf{G}_{12}^T & \mathbf{G}_{22} \end{bmatrix} = \left[\begin{array}{cc|cc} \mathbf{A} & \mathbf{0}_K & \mathbf{C} & \mathbf{D} \\ \mathbf{0}_K & \mathbf{A} & \mathbf{D}^T & -\mathbf{C}^T \\ \hline \mathbf{C}^T & \mathbf{D} & \mathbf{B} & \mathbf{0}_K \\ \mathbf{D}^T & -\mathbf{C} & \mathbf{0}_K & \mathbf{B} \end{array}\right],$$

where matrices $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$ and $\mathbf{D}$ are $K \times K$ circulant matrices. This can be seen from the fact that the matrices

$$\mathbf{A} = \sum_{i=1}^{2} \Big\{ \Re\{\mathbf{U}_i + \mathbf{V}_i\}\Re\{\mathbf{U}_i^H + \mathbf{V}_i^H\} - \Im\{\mathbf{V}_i - \mathbf{U}_i\}\Im\{\mathbf{V}_i^H - \mathbf{U}_i^H\} \Big\},$$

$$\mathbf{B} = \sum_{i=1}^{2} \Big\{ \Re\{\mathbf{U}_i - \mathbf{V}_i\}\Re\{\mathbf{U}_i^H - \mathbf{V}_i^H\} - \Im\{\mathbf{U}_i + \mathbf{V}_i\}\Im\{\mathbf{U}_i^H + \mathbf{V}_i^H\} \Big\},$$

$$\mathbf{C} = -\sum_{i=1}^{2} \Big\{ \Re\{\mathbf{U}_i + \mathbf{V}_i\}\Im\{\mathbf{U}_i^H + \mathbf{V}_i^H\} + \Im\{\mathbf{V}_i - \mathbf{U}_i\}\Re\{\mathbf{V}_i^H - \mathbf{U}_i^H\} \Big\},$$

$$\begin{aligned} \mathbf{D} = {} & \Re\{\mathbf{U}_2 + \mathbf{V}_2\}\Im\{\mathbf{U}_1 + \mathbf{V}_1\} - \Re\{\mathbf{U}_1 + \mathbf{V}_1\}\Im\{\mathbf{U}_2 + \mathbf{V}_2\} \\ & + \Im\{\mathbf{V}_2 - \mathbf{U}_2\}\Re\{\mathbf{V}_1 - \mathbf{U}_1\} - \Im\{\mathbf{V}_1 - \mathbf{U}_1\}\Re\{\mathbf{V}_2 - \mathbf{U}_2\}, \end{aligned}$$

are product and summation of circulant matrices. The Schur complement of $\mathbf{G}_{11}$ in $\mathbf{G}$ can be obtained as $\mathbf{G}_s = \mathbf{G}_{22} - \mathbf{G}_{12}^T \mathbf{G}_{11}^{-1} \mathbf{G}_{12} = \mathbf{I}_2 \otimes \mathbf{E}$, where $\mathbf{E} = \mathbf{B} - (\mathbf{C}^T \mathbf{A}^{-1} \mathbf{C} + \mathbf{D}\mathbf{A}^{-1}\mathbf{D}^T)$ is also a $K \times K$ circulant matrix. Hence only $\mathbf{A}^{-1}$, $\mathbf{B}^{-1}$ and $\mathbf{E}^{-1}$ are involved to invert matrices $(\mathbf{H}\mathbf{H}^T + \sigma_n^2/\sigma_s^2 \mathbf{I}_{4K})$ and $\mathbf{H}\mathbf{H}^T$. The circulant nature of matrices $\mathbf{A}$, $\mathbf{B}$ and $\mathbf{E}$ further indicates that their inversions can be easily calculated using unitary DFT matrix, i.e., $\mathbf{A}^{-1} = \mathbf{W}^H \mathbf{\Lambda}_a^{-1} \mathbf{W}$ with diagonal elements of $\mathbf{\Lambda}_a$ being DFT of the first column of $\mathbf{A}$.

The performance of the linear filter solutions can be further improved by introducing *nonlinear* filters, such as MMSE decision feedback (MMSE-DFE) and/or ZF-DFE, to remove the remaining inter-symbol interference (ISI). Let $\mathbf{F}_f = \mathbf{L}^T\mathbf{F}_{\text{LMMSE}}$ and $\mathbf{F}_b = \mathbf{L}^T - \mathbf{I}_{4K}$ denote the feedforward filter and feedback filter of MMSE-DFE, respectively, the $4K \times 4K$ lower-triangular matrix $\mathbf{L}$ can be explicitly obtained by the factorization, $\mathbf{H}^T\mathbf{H} + \sigma_n^2/\sigma_s^2\mathbf{I}_{4K} = \mathbf{L}\boldsymbol{\Lambda}_h\mathbf{L}^T$. Here $\boldsymbol{\Lambda}_h$ is a $4K \times 4K$ diagonal matrix used to normalize the diagonal entries of $\mathbf{L}$. The block symmetric matrix $\tilde{\mathbf{G}} = \mathbf{H}^T\mathbf{H}$, with the similar structure as $\mathbf{G}$, can be decomposed into

$$\tilde{\mathbf{G}} = \begin{bmatrix} \mathbf{I}_{2K} & \mathbf{0}_{2K} \\ \tilde{\mathbf{G}}_{12}^T\tilde{\mathbf{G}}_{11}^{-1} & \mathbf{I}_{2K} \end{bmatrix} \begin{bmatrix} \tilde{\mathbf{G}}_{11} & \mathbf{0}_{2K} \\ \mathbf{0}_{2K} & \tilde{\mathbf{G}}_s \end{bmatrix} \begin{bmatrix} \mathbf{I}_{2K} & \tilde{\mathbf{G}}_{11}^{-1}\tilde{\mathbf{G}}_{12} \\ \mathbf{0}_{2K} & \mathbf{I}_{2K} \end{bmatrix},$$

where $\tilde{\mathbf{G}}_{11} = \mathbf{I}_2 \otimes \tilde{\mathbf{A}}$, and $\tilde{\mathbf{G}}_s = \tilde{\mathbf{G}}_{22} - \tilde{\mathbf{G}}_{12}^T\tilde{\mathbf{G}}_{11}^{-1}\tilde{\mathbf{G}}_{12} = \mathbf{I}_2 \otimes \tilde{\mathbf{E}}$. Hence, obtaining matrix $\mathbf{L}$, only involves the inversion of $K \times K$ circulant matrix, whose computation load again can be dramatically reduced by DFT operations.

Dropping the diagonal loading factor, the implementation of ZF-DFE is straightforward.

B. Equivalent Frequency Domain Processing

By exploiting the structure of (5), a frequency domain processing can be applied to mitigate the I/Q imbalance. The equivalent frequency domain data model with white noise vector can be formulated as

$$\begin{bmatrix} \mathbf{W}\mathbf{r}_1 \\ \mathbf{W}\mathbf{J}\mathbf{r}_2^* \end{bmatrix} = \begin{bmatrix} \boldsymbol{\Lambda}_{u_1} & \boldsymbol{\Lambda}_{u_2} \\ \boldsymbol{\Lambda}_{u_2}^* & -\boldsymbol{\Lambda}_{u_1}^* \end{bmatrix} \begin{bmatrix} \mathbf{W}\mathbf{s}_1 \\ \mathbf{W}\mathbf{s}_2 \end{bmatrix} + \begin{bmatrix} \boldsymbol{\Lambda}_{v_1} & \boldsymbol{\Lambda}_{v_2} \\ \boldsymbol{\Lambda}_{v_2}^* & -\boldsymbol{\Lambda}_{v_1}^* \end{bmatrix} \begin{bmatrix} \mathbf{W}\mathbf{s}_1^* \\ \mathbf{W}\mathbf{s}_2^* \end{bmatrix} + \begin{bmatrix} \mathbf{W}\mathbf{n}_1 \\ \mathbf{W}\mathbf{J}\mathbf{n}_2^* \end{bmatrix}.$$

For K-point DFT, we have

$$\mathcal{F}\{\mathbf{s}_i[k]\} = \bar{\mathbf{s}}_i[k], \quad \mathcal{F}\{\mathbf{s}_i^*[k]\} = \bar{\mathbf{s}}_i^*[k'], \quad k = 0, \cdots, K-1,$$

where $\bar{\mathbf{s}}$ is DFT of $\mathbf{s}$, and $k' = (-k)_K$. Assume K is even, let $\bar{\mathbf{r}}_1 = \mathbf{W}\mathbf{r}_1$, $\underline{\bar{\mathbf{r}}}_2 = \mathbf{W}\mathbf{J}\mathbf{r}_2^*$, $\bar{\mathbf{n}}_1 = \mathbf{W}\mathbf{n}_1$, and $\underline{\bar{\mathbf{n}}}_2 = \mathbf{W}\mathbf{J}\mathbf{n}_2^*$. For $k = 0, K/2$, it can be obtained that

$$\underbrace{\begin{bmatrix} \bar{\mathbf{r}}_1[k] \\ \underline{\bar{\mathbf{r}}}_2[k] \end{bmatrix}}_{\mathbf{r}_a} = \underbrace{\begin{bmatrix} \lambda_{u_1}[k] & \lambda_{u_2}[k] \\ \lambda_{u_2}^*[k] & -\lambda_{u_1}^*[k] \end{bmatrix}}_{\mathbf{E}_u} \underbrace{\begin{bmatrix} \bar{\mathbf{s}}_1[k] \\ \bar{\mathbf{s}}_2[k] \end{bmatrix}}_{\mathbf{s}_a} + \underbrace{\begin{bmatrix} \lambda_{v_1}[k] & \lambda_{v_2}[k] \\ \lambda_{v_2}^*[k] & -\lambda_{v_1}^*[k] \end{bmatrix}}_{\mathbf{E}_v} \begin{bmatrix} \bar{\mathbf{s}}_1^*[k] \\ \bar{\mathbf{s}}_2^*[k] \end{bmatrix} + \underbrace{\begin{bmatrix} \bar{\mathbf{n}}_1[k] \\ \underline{\bar{\mathbf{n}}}_2[k] \end{bmatrix}}_{\mathbf{n}_a}, \tag{8}$$

where $\lambda_{u_m}[k]$ and $\lambda_{v_m}[k]$ are the k-th diagonal entries of $\boldsymbol{\Lambda}_{u_m}$ and $\boldsymbol{\Lambda}_{v_m}$, respectively. Reformulate (8) as

$$\underbrace{\begin{bmatrix} \Re\{\mathbf{r}_a\} \\ \Im\{\mathbf{r}_a\} \end{bmatrix}}_{\bar{\mathbf{r}}_a} = \underbrace{\begin{bmatrix} \Re\{\mathbf{E}_u + \mathbf{E}_v\} & \Im\{\mathbf{E}_v - \mathbf{E}_u\} \\ \Im\{\mathbf{E}_u + \mathbf{E}_v\} & \Re\{\mathbf{E}_u - \mathbf{E}_v\} \end{bmatrix}}_{\mathbf{H}_a} \underbrace{\begin{bmatrix} \Re\{\mathbf{s}_a\} \\ \Im\{\mathbf{s}_a\} \end{bmatrix}}_{\bar{\mathbf{s}}_a} + \begin{bmatrix} \Re\{\mathbf{n}_a\} \\ \Im\{\mathbf{n}_a\} \end{bmatrix},$$

then the least-squares (LS) estimate of $\bar{\mathbf{s}}_a$ can be obtained as

$$\hat{\bar{\mathbf{s}}}_a = (\mathbf{H}_a^T\mathbf{H}_a)^{-1}\mathbf{H}_a^T\bar{\mathbf{r}}_a.$$

For other values of k, it can be obtained that

$$\underbrace{\begin{bmatrix} \bar{\mathbf{r}}_1[k] \\ \underline{\bar{\mathbf{r}}}_2[k] \\ \bar{\mathbf{r}}_1^*[k'] \\ \underline{\bar{\mathbf{r}}}_2^*[k'] \end{bmatrix}}_{\bar{\mathbf{r}}_b} = \underbrace{\begin{bmatrix} \lambda_{u_1}[k] & \lambda_{u_2}[k] & \lambda_{v_1}[k] & \lambda_{v_2}[k] \\ \lambda_{u_2}^*[k] & -\lambda_{u_1}^*[k] & \lambda_{v_2}^*[k] & -\lambda_{v_1}^*[k] \\ \lambda_{v_1}^*[k'] & \lambda_{v_2}^*[k'] & \lambda_{u_1}^*[k'] & \lambda_{u_2}^*[k'] \\ \lambda_{v_2}[k'] & -\lambda_{v_1}[k'] & \lambda_{u_2}[k'] & -\lambda_{u_1}[k'] \end{bmatrix}}_{\mathbf{H}_b} \underbrace{\begin{bmatrix} \bar{\mathbf{s}}_1[k] \\ \bar{\mathbf{s}}_2[k] \\ \bar{\mathbf{s}}_1^*[k'] \\ \bar{\mathbf{s}}_2^*[k'] \end{bmatrix}}_{\bar{\mathbf{s}}_b} + \begin{bmatrix} \bar{\mathbf{n}}_1[k] \\ \underline{\bar{\mathbf{n}}}_2[k] \\ \bar{\mathbf{n}}_1^*[k'] \\ \underline{\bar{\mathbf{n}}}_2^*[k'] \end{bmatrix}.$$

Then the LS estimate of $\bar{\mathbf{s}}_b$ can be obtained as

$$\hat{\bar{\mathbf{s}}}_b = (\mathbf{H}_b^H\mathbf{H}_b)^{-1}\mathbf{H}_b^H\bar{\mathbf{r}}_b.$$

It shows that $(K/2+1)$ inversions of 4×4 matrix is needed in frequency domain to obtain the LS estimate. In addition, due to the special structure of matrix $\mathbf{H}_a$ and $\mathbf{H}_b$, by using the matrix inversion formula, all the sub-matrices to be inverted to obtain $(\mathbf{H}_a^T\mathbf{H}_a)^{-1}$ and $(\mathbf{H}_b^H\mathbf{H}_b)^{-1}$ are diagonal [10], reducing the computational complexity further.

The LS estimate of $\bar{\mathbf{s}}_i$, denoted as $\hat{\bar{\mathbf{s}}}_i$, can be obtained by stacking the elements of $\hat{\bar{\mathbf{s}}}_a$ and $\hat{\bar{\mathbf{s}}}_b$ correspondingly. Then LS estimate of $\mathbf{s}_i$ can be obtained from

$$\hat{\mathbf{s}}_{i,LS} = \mathbf{W}^H\hat{\bar{\mathbf{s}}}_i. \tag{9}$$

C. Estimation of Channel and I/Q Imbalance

Since only ECSI such as $\mathbf{u}_i$ and $\mathbf{v}_i$ are needed in the proposed solutions, the tasks of separately estimating the channel parameters $\mathbf{h}_i$ as well as the I/Q imbalance parameters $\{\alpha_T, \beta_T, \alpha_R, \beta_R\}$ can be combined [8]. Here the LS method is given to demonstrate the feasibility.

During the training period, according to (3) (5), we have

$$\underbrace{\begin{bmatrix} \mathbf{r}_{t,1} \\ \mathbf{r}_{t,2} \end{bmatrix}}_{\mathbf{r}_t} = \underbrace{\begin{bmatrix} \mathbf{S}_{t,1} & \mathbf{S}_{t,2} & \mathbf{S}_{t,1}^* & \mathbf{S}_{t,2}^* \\ -\mathbf{S}_{t,2}^* & \mathbf{S}_{t,1}^* & -\mathbf{S}_{t,2} & \mathbf{S}_{t,1} \end{bmatrix}}_{\mathbf{S}_t} \underbrace{\begin{bmatrix} \mathbf{u}_1 \\ \mathbf{u}_2 \\ \mathbf{v}_1 \\ \mathbf{v}_2 \end{bmatrix}}_{\boldsymbol{\theta}} + \begin{bmatrix} \mathbf{n}_{t,1} \\ \mathbf{n}_{t,2} \end{bmatrix}, \tag{10}$$

where the $N_t \times (L+1)$ matrix $\mathbf{S}_{t,i}$ is the first $L+1$ columns of a $N_t \times N_t$ circulant matrix $\text{circ}(\mathbf{s}_{t,i})$, with $\mathbf{s}_{t,i}$ being the $N_t \times 1$ training data vector sending from the i^{th} antenna. Consequently, the estimated ECSI can be obtained as

$$\hat{\boldsymbol{\theta}} = (\mathbf{S}_t^H\mathbf{S}_t)^{-1}\mathbf{S}_t^H\mathbf{r}_t. \tag{11}$$

To study the performance degradation caused by I/Q imbalance without compensation, only $\mathbf{u}_i$ is estimated, then $\boldsymbol{\theta}$ and $\mathbf{S}_t$ in (10) become

$$\tilde{\boldsymbol{\theta}} = \begin{bmatrix} \mathbf{u}_1 \\ \mathbf{u}_2 \end{bmatrix}, \qquad \tilde{\mathbf{S}}_t = \begin{bmatrix} \mathbf{S}_{t,1} & \mathbf{S}_{t,2} \\ -\mathbf{S}_{t,2}^* & \mathbf{S}_{t,1}^* \end{bmatrix}.$$

Once $\hat{\tilde{\boldsymbol{\theta}}}$ is available, the conventional detection for TR-OSTBC system is performed.

The extension of our solution to the general multiple-input multiple-output (MIMO) systems is straightforward.

IV. Simulation Experiments

In our simulations, L-order channels with $\mathbf{h}_i \sim \mathcal{CN}(\mathbf{0}, \mathbf{I}_{L+1})$, data block length $K = 20$, CP length $L = 2$ and training data vector length $N_t = \{10, 20, 40\}$ are used. In addition, SNR $\triangleq \sigma_s^2/\sigma_n^2$, and the power loss due to CP is neglected because of the fact $K + L \gg L$.

Results in Fig.1∼3 demonstrate the system performance, SER as a function of SNR, of the proposed solutions. As can be seen that the I/Q imbalance causes dramatic performance degradation without compensation, and receiver I/Q imbalance causes more performance degradation than transmitter I/Q imbalance in such a system. In addition, the equivalent frequency domain LS (FD-LS) method provides the same performance as the time domain ZF method. The LMMSE method improves

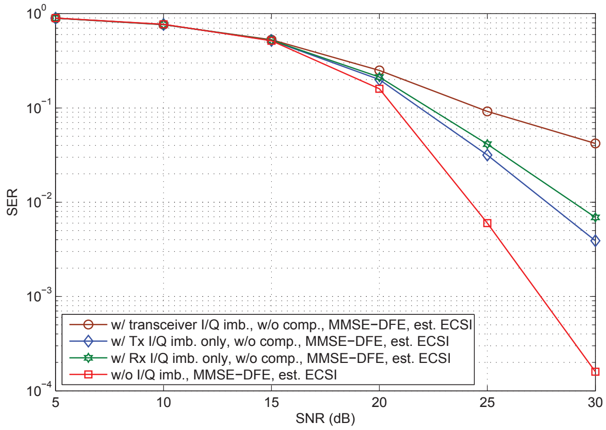

Fig. 1. SER versus SNR of a 2×1 TR-OSTBC system with I/Q imbalance. System parameters are: $\epsilon_T = 5\%$, $\varphi_T = 5°$, $\epsilon_R = 5\%$, $\varphi_R = 5°$, $K = 20$, $L = 2$, and $N_t = 20$. A 64-QAM constellation is used. Number of independent trials is 10^6.

the performance compared to the above two. Moreover, further performance improvement can be achieved by introducing non-linear ZF-DFE and MMSE-DFE, compared to their linear counterparts. Furthermore, Fig.4 shows the proposed solutions with estimated ECSI can mitigate the I/Q imbalance as well.

V. Conclusions

This work develops the solutions to symbol detection, both in time domain and in frequency domain, for TR-STBC systems with I/Q imbalance operating over frequency-selective fading channels. Results demonstrate the effectiveness of the proposed approaches in mitigating I/Q imbalance and improving the SER performance.

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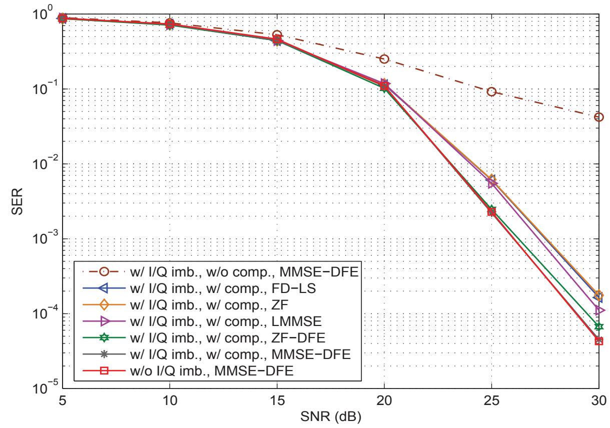

Fig. 2. SER versus SNR of a 2×1 TR-OSTBC system with I/Q imbalance. System parameters are: $\epsilon_T = 5\%$, $\varphi_T = 5°$, $\epsilon_R = 5\%$, $\varphi_R = 5°$, $K = 20$, and $L = 2$. A 64-QAM constellation is used, and ECSI is assumed known. Number of independent trials is 10^6.

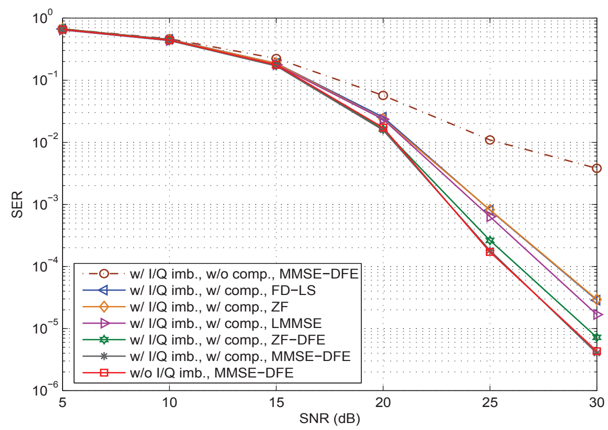

Fig. 3. SER versus SNR of a 2×1 TR-OSTBC system with I/Q imbalance. System parameters are: $\epsilon_T = 5\%$, $\varphi_T = 5°$, $\epsilon_R = 5\%$, $\varphi_R = 5°$, $K = 20$, and $L = 2$. A 16-PSK constellation is used, and ECSI is assumed known. Number of independent trials is 10^6.

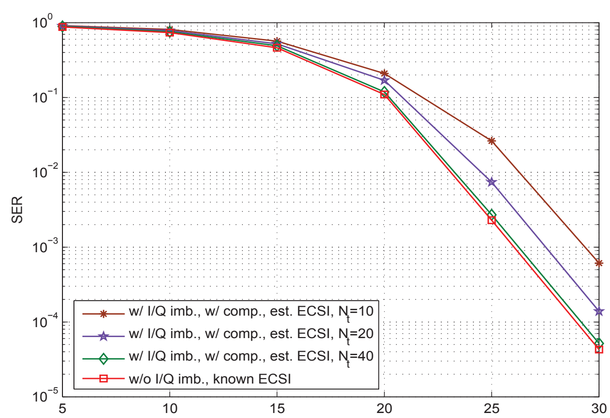

Fig. 4. SER versus SNR of a 2×1 TR-OSTBC system with I/Q imbalance. System parameters are: $\epsilon_T = 5\%$, $\varphi_T = 5°$, $\epsilon_R = 5\%$, $\varphi_R = 5°$, $K = 20$, $L = 2$, and $N_t = \{10, 20, 40\}$. A 64-QAM constellation is used, and MMSE-DFE is performed. Number of independent trials is 10^6.